

Broadband Planar 5:1 Impedance Transformer

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Abstract—This paper presents a broadband Guanella-type planar impedance transformer that transforms $50\ \Omega$ to $10\ \Omega$ with a 10 dB bandwidth of 1–14 GHz. The transformer is designed on a flexible $50\ \mu\text{m}$ thick polyimide substrate in microstrip and parallel-plate transmission line topologies, and is inspired by the traditional 4:1 Guanella transformer. Back-to-back transformers were designed and fabricated for characterization in a $50\ \Omega$ system. Simulated and measured results are in excellent agreement.

Index Terms—Transformers, impedance matching, broadband, parallel-plate line.

I. INTRODUCTION

TRANSMISSION line transformers (TLTs) are widely used as impedance matching networks in radio frequency applications [1]–[14]. TLTs, first implemented by Guanella in 1944 [2], can simultaneously exhibit an octave of bandwidth for discrete impedance transformation ratios, are compact, and attractive for use as broadband matching networks. A 4:1 Guanella-type TLT nominally exhibits frequency-independent characteristics when realized with a pair of appropriate impedance and equal-delay transmission lines. Fig. 1 (a) shows the schematic implementation of a 4:1 impedance transformer, where two delay lines of equal length are connected such that currents add in phase at the low-impedance end. As a result of the delay and line symmetry, the transformation is theoretically independent of line length at finite frequencies. A simpler and more common type of TLT is the Ruthroff transformer [3], which appears similar to the Guanella “equal delay” design, but differs in implementation with its use of a single transmission line delay. As a result, the Ruthroff transformer has a smaller footprint, however, its response is not frequency independent and its bandwidth is ultimately limited by the transmission line length.

Guanella TLT performance is a function of the transmission line impedance, delay, and symmetry. At UHF and low microwave frequencies, TLTs are implemented with coaxial lines; in order to decrease their low-frequency limit and suppress unbalanced currents, they are wound around ferrite cores [4]–[6]. The high-frequency response is limited by the transmission line interconnect junction parasitics; to avoid this limitation, compensation of the junctions and more precise fabrication is required to ensure suppression and control over parasitic reactances. A Guanella-type 4:1 transformer in wafer-scale micro-coaxial technology with an operating bandwidth of 2–24 GHz was recently implemented [7].

Most prior published planar TLTs are limited to the Ruthroff configuration. Examples have been implemented in monolithic

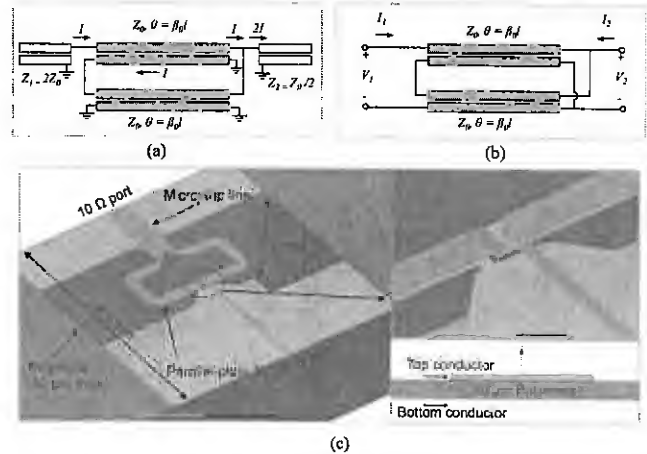


Fig. 1. (a) Circuit model of a conventional 4:1 Guanella impedance transformer. (b) Generalized circuit model of the two-transmission lines Guanella impedance transformer. (c) Rendered image of the transformer over a $10\ \text{mm} \times 4.5\ \text{mm} \times 4\ \text{mm}$ cavity. The input and output transmission lines are microstrip for ease of integration with standard printed circuit board connectors. The bottom right inset shows the parallel-plate transmission line topology with one conductor on top and one conductor on the bottom used to implement the 5:1 planar transformer. The top right inset shows the connection at the high impedance (Z_2) end of the transformer; the parasitics associated with this series connection limit the high-frequency performance of the transformer.

microwave integrated circuits [4], [8], [9], several were implemented with coupled microstrip lines [10], [11], and one recent example for superconducting applications was implemented as a $6.25\ \Omega$ to $25\ \Omega$ transformer from 2–13 GHz [12].

In this paper we describe a planar Guanella-type 5:1 TLT that is easily integrable with common planar transmission line topologies, *e.g.*, microstrip. A primary goal of this study is to demonstrate a compact planar broadband matching circuit that is both cost effective and easily implementable with printed circuit boards (PCBs). Due to use of a durable and flexible substrate, this transformer can be used in extreme environments, including at cryogenic temperatures.

II. TRANSFORMER DESIGN PROCEDURE

A 4:1 Guanella transformer (Fig. 1(a)) consists of two equal-length, equal characteristic impedance (Z_0) transmission lines that are connected in series at the high-impedance end (Z_1) and in parallel at the low-impedance end (Z_2). The impedance transformation is depicted in Fig. 1(a); current flows in both conductors of a transmission line but in opposite directions. Starting from the high-impedance end, current I flows in the top conductor of the first transmission line (Z_0); an equal and opposite current flows in the bottom conductor of the same line. By adding a second transmission line with the same length and characteristic impedance, and then connecting the

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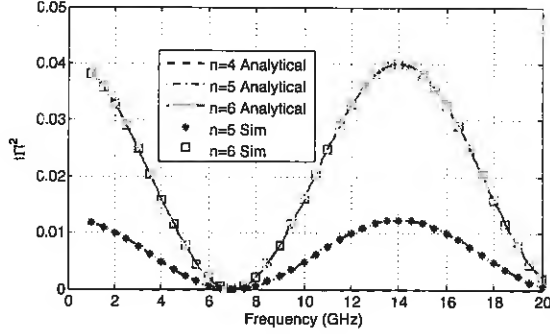


Fig. 2. Calculated and simulated reflection coefficient as a function of frequency for impedance ratios $n = 4, 5$ and 6 for the two-transmission line impedance transformer shown in Fig. 1 (b), where $Z_0 = (Z_1 Z_2)^{1/2}$, and $l = 10.714$ mm ($\lambda/4$ at 7 GHz in free-space). Simulation is performed using ideal circuit model in a circuit simulator.

two transmission lines in parallel and series at each end, a 4:1 impedance transformation ratio is achieved. The relationship between characteristic impedances is $Z_0 = (Z_1 Z_2)^{1/2}$ [1]. Due to the transmission lines' equal length and impedance, the differential mode is theoretically frequency independent. In practice, the structure's port isolation and match are inter-related, and the low frequency response is set by the length of the two transmission lines [14]. The high-frequency limit is optimized by minimizing the junction parasitics at the series and parallel transmission line connections.

A. 5:1 Impedance Transformer Design

A Guanella TLT can be realized for a discrete transformation ratio where the square root of the desired ratio n is equal to a rational number. If this condition is met, the ratio can be realized by connecting multiple transmission lines in series and parallel [1]. Therefore, in its simplest form, the Guanella topology does not support a 5:1 ratio. The closest realizable impedance ratios are $(\frac{7}{3})^2:1$ or $(\frac{11}{5})^2:1$, which are attainable with 5 and 7 transmission lines, respectively. The transmission lines are connected in parallel and series combination at the input and output. The use of multiple lines requires a larger footprint and more complex design with additional parasitics from multiple connections, which reduces the bandwidth and creates more resonances in the passband.

For a simplified realization of a 5:1 ratio, we used a similar topology as Guanella 4:1 transformer (Fig. 1 (b)). For the 5:1 transformer, due to deviation from the ideal 4:1 case, there is a length dependency where the phase cancels out completely at $l = \lambda/4$. In the absence of parasitics the odd mode input impedance looking into port 1 can be calculated using (1) [13]. The reflection coefficient Γ is calculated by (2). Fig. 2 shows the transmission line calculation and simulated $|\Gamma|^2$ as a function of frequency for ratios $n = 4, 5$, and 6 , where $Z_0 = (Z_1 Z_2)^{1/2}$ and $l = 10.714$ mm ($\lambda/4$ at 7 GHz in free space).

$$Z_{in} = 2Z_0 \frac{2Z_2 + jZ_0 \tan(\beta l)}{Z_0 + 2jZ_2 \tan(\beta l)} \quad (1)$$

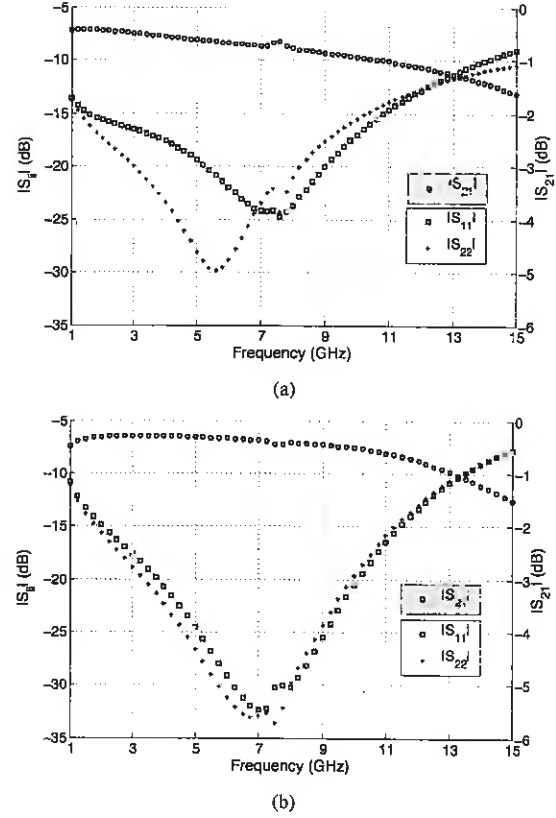


Fig. 3. (a) Simulated S-parameter results of the transformer in free space. (b) Simulated S-parameter results of the transformer above the 10 mm \times 4.5 mm \times 4 mm cavity. The transmission lines' phases cancel at 7 GHz ($l = \lambda/4$) resulting in the best match.

$$\Gamma = \frac{Z_{in} - Z_1}{Z_{in} + Z_1} \quad (2)$$

To implement the transformer in a planar topology, we chose a parallel-plate transmission line with one conductor atop the substrate and one conductor beneath the substrate for the two equal delays. The substrate is 50 μ m thick polyimide with $\epsilon_r = 3.4$ and $\tan \delta = 2 \times 10^{-3}$ and the design impedances are $Z_0 = 22.3 \Omega$, $Z_1 = 10 \Omega$, and $Z_2 = 50 \Omega$. The input and output transmission lines are microstrip, so that they can easily be integrated with other components on a PCB. Fig. 1 (c) shows a 3D rendering of the design above a 10 mm \times 4.5 mm \times 4 mm cavity. In this design a 10 Ω microstrip line transitions to a 10 Ω parallel-plate transmission line, then connects in parallel to a pair of 5.5 mm long 22 Ω parallel-plate transmission lines. The parallel-plate transmission lines then connect in series with two 75 μ m diameter vias (Fig. 1 (c) inset), and finally connect to a 50 Ω microstrip line.

The transformer was first electromagnetically modeled using the finite element method (FEM) in an air-boundary box, Fig. 3 (a) shows the simulated S-parameter results of this model. This model has return loss better than 10 dB from 1–14 GHz with less than 1 dB transmission loss at 10 GHz.

Since the transformer needs to be packaged and integrated with external components, it is designed above a metallic 10 mm \times 4.5 mm \times 4 mm cavity. The cavity is required since the transformer cannot be attached to a conductive mount,

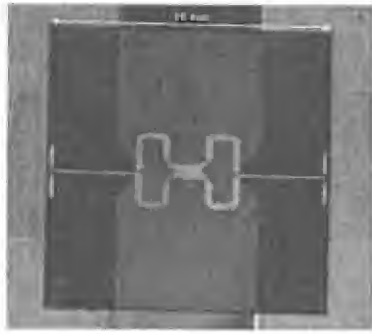


Fig. 4. Photograph of the fabricated back-to-back impedance transformer on a 50 μm thick polyimide substrate. The transformer was placed above the ground to prevent shorting of the bottom conductor of the parallel-plate transmission line.

as it would short the transformer. Additionally, because the transformer transmission lines are on a low dielectric constant material, the fringing fields at the edges of the conductors are significant and as a result there is significant radiation loss. By designing the transformer above an open cavity, the fields are confined and the overall loss of the device is reduced from 1 dB to 0.5 dB at 10 GHz. Further reductions are possible by enclosing the entire transmission line circuit in a non-resonant cavity. Fig. 3 (b) shows the simulated S-parameter results of the transformer placed above a cavity. The small resonant feature that appears at 7.5 GHz in both simulations is due to a small length difference of the two main transmission lines caused by the series interconnection at the high-impedance end. This phenomena creates a path for a weakly-coupled $\lambda/2$ shorted resonator within the transformer, allowing a small resonance feature to appear [7].

III. CHARACTERIZATION

To characterize the transformer in a 50 Ω system, we designed and fabricated a symmetric structure in which two transformers are connected at the 10 Ω port. Fig. 4 is a photograph of the back-to-back transformers recessed above a ground plane for characterization. A custom through-reflect-line (TFL) calibration set with two lines, an open, and a thru was used. Fig. 5 shows the simulated and measured S-parameter results of the back-to-back transformers over the range of 1–12 GHz. The simulation and measured results show very good agreement. The slight discrepancy between the measured and simulated $|S_{11}|$ is due to a manufacturing error; some of the transmission lines had small errors in their fabricated widths, resulting in slightly different characteristic impedances than specified in the design.

IV. CONCLUSION

In this paper we presented a planar Guanella-type impedance transformer with a ratio of 5:1. Based on these results, it is possible to employ the Guanella transformer topology beyond the 5:1 ratio presented here. As we deviate from the ideal 4:1 ratio in a Guanella-type transformer, the return loss will be degraded. In practice, the achievable upper frequency response is limited by the parasitics' reactance (e.g.,

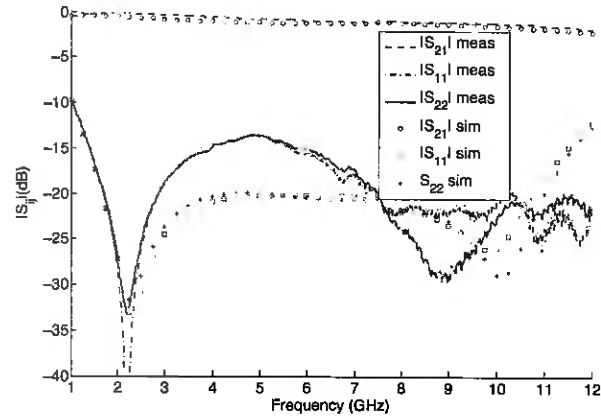


Fig. 5. Measured and simulated S-parameter results of the back-to-back transformers. The transformer is positioned above the ground plane to prevent the circuit from shorting. The phase cancellation (resonance) at 2.2 GHz and 9 GHz is due to the back-to-back measurement configuration.

via inductance) and ones ability to compensate for these reactances. A Guanella-type transformer has been implemented in a planar circuit topology, using easily-fabricated transmission lines with two metallization layers. Some of the benefits of this transformer are its high impedance transformation ratio, compact size (1 cm \times 1 cm), low cost, and durability in extreme environments due to utilization of a flexible, thin, polyimide substrate. Compared to conventional broadband matching circuits such as tapers, they are an order of magnitude smaller in electrical length, resulting in less dielectric and metal loss.

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